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ORTHOGONAL RE-SPREAD FOR UPLINK WCDMA BEAMFORMING

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Abstract - This paper presents a new extraction method for the spatial correlation matrix of interference and noise only based on orthogonal re-spreading. The method uses one of the remaining orthogonal variable spreading factor (OVSF) code that is orthogonal to all traffic and control channels to spread energy from the desired user while attempting to maintain the same level of co-channel interference. De-spreading with the assigned OVSF codes that is performed in parallel provides the usual noisy correlation matrix. With those two matrices at hand optimal beamforming using maximal signal-to-interference ratio (MSIR) criterion is readily performed. Although the matrix extracted by the orthogonal re-spread is not a statistically consistent estimate, remarkable potentials in practical situations are demonstrated.

I. INTRODUCTION

Adaptive antennas are widely recognised as one of the key technologies to meet an immense demand for spectral efficiency of the third generation mobile communications systems. The 3G proposals are designed in such a way to facilitate the adaptive antenna concepts at the standardisation level e.g. by employing connection dedicated pilot sequences on both uplink and downlink as in WCDMA [8].

Many DS-CDMA receiver architectures with adaptive antennas have been investigated recently - [2,3,4,5] amongst others. The architectures range in complexity and performance. In [2] simple space-time processing scheme is proposed, where a despreader follows optimal processing in space domain. This approach can be extended to 2D-rake receiver with maximal ratio combining in the time domain. Joint space-time domain equalisation without neither detailed channel knowledge nor training sequences has been investigated in [3]. Reduced complexity methods are presented in [4,5] where in principle signal and noisy time bins are treated separately.

All of the aforementioned approaches rely on extraction of the space-time correlation matrices of interference and

noise only \mathbf{R}_{IN} . The first approach [2] (later adopted also in [3]) extracts \mathbf{R}_{IN} as a matrix of difference between pre-processed antenna outputs and the scaled matched filter outputs. This method however may suffer from the signal leakage, which is due to the fact that even before de-spreading the desired signal can have significant magnitude. More computationally demanding approaches are presented in [4] and [5] where \mathbf{R}_{IN} is estimated in the chip-shifted time-bins when multipaths of the desired user are not present. Those methods however involve an additional test for the signal presence in the time bins.

In what follows we propose another method that estimates \mathbf{R}_{IN} with minimal complexity. The presented method is suitable for DS-CDMA systems with aperiodic spreading sequences where the orthogonal codes used for signal spreading are masked by the cell specific scrambling codes as in WCDMA. In that sense the method directly exploits the uplink structure of the WCDMA. The training sequences however are not essential as the signal vectors can be estimated blindly.

This paper is organised as follows: The next section briefly discusses choices of the cost functions for the beamformer weight vector. Section III presents the method for the extraction of the correlation matrices and the uplink structure of the WCDMA. Application of the principal component method in the case where training sequences are not available is considered in section IV. Numerical examples and conclusions that follow wrap up this paper.

II. OPTIMAL CRITERIA

The beamformer weights can be chosen to directly maximise the signal-to-interference ratio. If the correlation matrices of signal only (\mathbf{R}_{SS}) and interference and noise only (\mathbf{R}_{IN}) are known, then the problem can be written as:

$$\mathbf{w}_{MSIR} = \arg \max_{\mathbf{w}} \left\{ \frac{\mathbf{w}^H \mathbf{R}_{SS} \mathbf{w}}{\mathbf{w}^H \mathbf{R}_{IN} \mathbf{w}} \right\} \quad (1)$$

Where: \mathbf{w} is the beamformer weight vector.
The solution to (1) is given by [7]:

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$$\mathbf{R}_{SS} \mathbf{w} = \frac{\mathbf{w}^H \mathbf{R}_{SS} \mathbf{w}}{\mathbf{w}^H \mathbf{R}_{IN} \mathbf{w}} \mathbf{R}_{IN} \mathbf{w} \quad (2)$$

Which is a joint eigenvalue problem ($\mathbf{R}_{SS}, \mathbf{R}_{IN}$). Since \mathbf{R}_{IN} is always strictly positive definite (due to ubiquitous noise), (2) can be written:

$$\mathbf{R}_{IN}^{-1} \mathbf{R}_{SS} \mathbf{w} = \lambda \mathbf{w} \quad (3)$$

Where: $\lambda = \frac{\mathbf{w}^H \mathbf{R}_{SS} \mathbf{w}}{\mathbf{w}^H \mathbf{R}_{IN} \mathbf{w}}$ is an eigenvalue of the standard

eigenvalue problem - (3). Associated with the maximal eigenvalue ($\lambda_{\max} = SIR_{\max}$) is the eigenvector \mathbf{w}_{MSIR} , which represents the optimum beamformer weights. If the spreading factor (SF) is large, then the effect of interchip interference (ICI) can be neglected [4]. In that case the signal correlation matrix is rank one - $rank\{\mathbf{R}_{SS}\}=1$ due to the presence of coherent multipaths. Eq. (3) then can be further reduced to:

$$\mathbf{w}_{MSIR} = \beta \mathbf{R}_{IN}^{-1} \mathbf{v} \quad (4)$$

where: \mathbf{v} is the crosscorrelation vector between received and desired signal and β a constant.

Another popular criterion for statistically optimum beamforming is *Minimum Mean Squared Error* with a solution given by the Wiener-Hopf equation:

$$\mathbf{w}_{MMSE} = \arg \min_{\mathbf{w}} E\{(\mathbf{d} - \mathbf{w}^H \mathbf{x})^2\} = \mathbf{R}_{XX}^{-1} \mathbf{v} \quad (5)$$

Where: \mathbf{d} and \mathbf{x} – is the desired and received signals respectively. The virtue of the MMSE solution is that it does not require the two aforementioned matrices (e.g. \mathbf{R}_{SS} and \mathbf{R}_{IN}). The only essential information is the correlation matrix of the observed signal \mathbf{R}_{XX} and cross-correlation vector with the desired signal.

With the *Minimum Variance Distortionless Response* MVDR criterion an objective is to minimise the output energy subject to the desired signal remains unchanged. Since in principle MVDR and MMSE share the solution (5) we will concentrate on MMSE and MSIR only.

It has been shown in [6] and later for aperiodic spreading codes in [4] that asymptotically the two solutions converge to the same vector. This is also evident from the Woodbury's identity [7]:

$$\mathbf{R}_{XX}^{-1} = \left[\frac{1}{1 + E\{\mathbf{s}^2\} \mathbf{v}^H \mathbf{R}_{IN}^{-1} \mathbf{v}} \right] \mathbf{R}_{IN}^{-1} \quad (6)$$

Which means that \mathbf{R}_{XX} and \mathbf{R}_{IN} are identical up to a real constant given by the expression in the brackets.

Nevertheless in practice estimates collected over N_s realisations are only available. Those estimates may not be very accurate and the error will cause the performance degradation. The result is the desired signal cancellation due to the mismatch between the presumed signal vector and the actual vector. This is particularly evident in the MMSE where poorly estimated signal vector is interpreted as interference. Signal cancellation problems in adaptive arrays are well-known (see [1] and the references therein).

For this reason the MSIR is preferable in practice if only extraction of \mathbf{R}_{IN} is possible.

III. EXTRACTION OF THE CORRELATION MATRICES

The method presented here exploits the structure of the uplink WCDMA shown in figure 1.

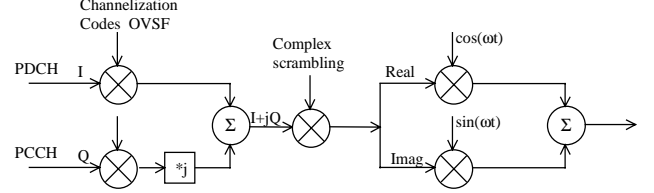


Figure 1: Structure of the uplink W-CDMA [8].

The physical channels PDCH (data) and PCCH (control) are mapped to I and Q branches respectively. Both branches are then spread by two different orthogonal variable spreading factor (OVSF) channelisation codes and scrambled by the complex code. Each part of the complex scrambling code is either long Gold code (40960 chips) or short Kasami code (256 chips).

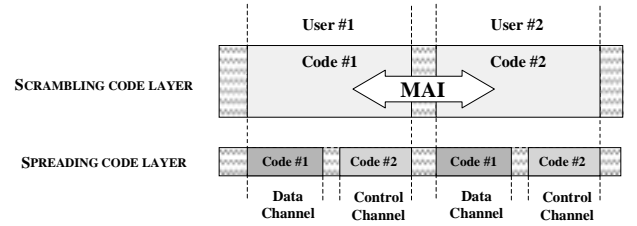


Figure 2: De-scrambling and de-spreading at the receiver of W-CDMA system (uplink).

At the receiver site the two operations of de-scrambling and de-spreading can be depicted in the layered schematic shown in figure 2. The scrambling codes are used to maintain semi-orthogonality for all possible lags between users in the asynchronous system.

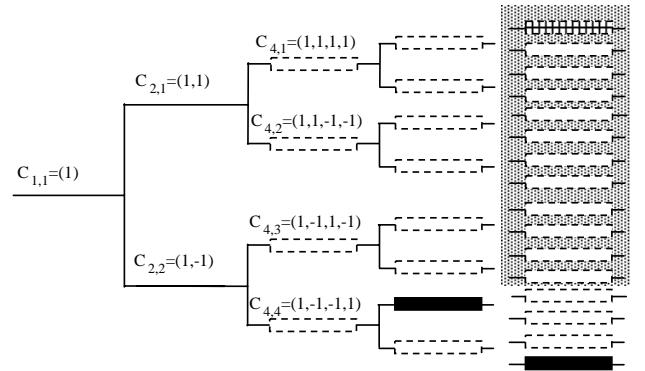


Figure 3: OVSF codes tree; PDCH and PCCH codes marked black; choice of codes for the orthogonal re-spread – shaded.

This semi-orthogonality is also a source of multiple access interference (MAI). After de-scrambling both data and control channel remain mutually orthogonal, which is guaranteed by OVFS codes and the fact that they are transmitted in the same physical radio channel.

The proposed method uses one of the remaining OVFS code that is orthogonal to all traffic and control channels. De-spreading with such a code (termed here orthogonal re-spreading) ensures that all control and data channels of the desired user are removed.

In general each OVFS code will produce different MAI. However long scrambling codes prevent the same MAI realisation in consecutive data symbols with one particular OVFS code. In these circumstances all codes are equally good. To cut down on processing burden it is proposed to use the first OVFS code, where the re-spreading is equivalent to summation over the symbol period.

In mathematical terms the whole reception process described above can be summarised with the following signal model:

$$\mathbf{x}(t) = \sum_{i=1}^M \sum_{l=1}^L \sqrt{p_{i,l}} z_{i,l}(t - \tau_{i,l}) s(t - \tau_{i,l}) \mathbf{a}_{i,l} + \mathbf{n}(t) \quad (7)$$

Where: \mathbf{x} represents received signal vector by the N element antenna array, \mathbf{a} is the $(N \times 1)$ antenna response vector, z – scrambling code, L – number of multipath components, M – number of users, p – power, and s is given by:

$$s(t - \tau_{i,l}) = c_{i,l}^{(d)}(t - \tau_{i,l}) b_{i,l}^{(d)}(t - \tau_{i,l}) + j \cdot c_{i,l}^{(c)}(t - \tau_{i,l}) b_{i,l}^{(c)}(t - \tau_{i,l}) \quad (8)$$

Where: $c^{(d)}$ is OVFS code assigned to the PDCH (data channel), $b^{(d)}$ is the PDCH sequence, $c^{(c)}$ is OVFS code assigned to the PCCH (control channel), $b^{(c)}$ is the PDCH sequence.

The received de-spread PCCH signal is given by:

$$\mathbf{y}_D(t) = -j \cdot \int_{\tau}^{T+\tau} z_{i,l}(t - \tau_{i,l}) c_{i,l}^{(c)}(t - \tau_{i,l}) \mathbf{x}(t) dt \quad (9)$$

Applying re-spread code $c^{(r)}$ which is orthogonal to both PDCH and PCCH channels:

$$c^{(d)}(c^{(r)})^T = c^{(c)}(c^{(r)})^T = 0 \quad (10)$$

The re-spread signal is given by:

$$\mathbf{y}_R(t) = -j \cdot \int_{\tau}^{T+\tau} z_{i,l}(t - \tau_{i,l}) c_{i,l}^{(r)}(t - \tau_{i,l}) \mathbf{x}(t) dt \quad (11)$$

Now de-spread and re-spread sample correlation matrices can be defined as:

$$\hat{\mathbf{R}}_{XX} = \frac{1}{N_S} \sum_{i=1}^{N_S} \mathbf{y}_D(t) \mathbf{y}_D^H(t) \quad (12)$$

$$\hat{\mathbf{R}}_{IN} = \frac{1}{N_S} \sum_{i=1}^{N_S} \mathbf{y}_R(t) \mathbf{y}_R^H(t) \quad (13)$$

Figure 4 depicts example for the orthogonal re-spread. As can be seen Capon azimuthal estimate for the three interfering sources is nearly the same for both de-spread and re-spread signals, whereas the desired signal is entirely suppressed in the re-spread case.

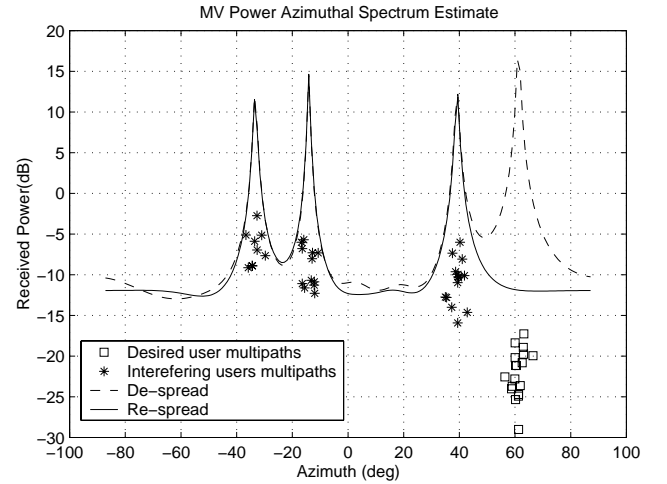


Figure 4: The desired signal suppression via the orthogonal re-spread.

IV. BLIND ADAPTATION – PRINCIPAL COMPONENT METHOD

In this section we assume there is no training sequence available in the system – in that sense the method is blind. The basic assumption for the principal component method developed in [2] is that the signal correlation matrix is rank one. Consequently \mathbf{R}_{SS} is an outer product of the signal vector. However in general the noise component and the estimation errors particularly in $\hat{\mathbf{R}}_{IN}$ will make the following matrix full rank:

$$\hat{\mathbf{R}}_{SS} = \hat{\mathbf{R}}_{XX} - \hat{\mathbf{R}}_{IN} \quad (14)$$

As such we have to find such a matrix $\tilde{\mathbf{R}}_{SS}$ that fits best $\hat{\mathbf{R}}_{SS}$ in the least squares sense, subject to condition - $\text{rank}[\tilde{\mathbf{R}}_{SS}] = 1$, which can be stated:

$$\tilde{\mathbf{R}}_{SS} = \arg \min_{\tilde{\mathbf{R}}_{SS} \text{ such rank}[\tilde{\mathbf{R}}_{SS}] = 1} \|\hat{\mathbf{R}}_{SS} - \tilde{\mathbf{R}}_{SS}\|_F^2 \quad (15)$$

This is a well known problem of low-rank modelling [7] with a solution:

$$\tilde{\mathbf{R}}_{SS} = \tilde{\mathbf{v}} \tilde{\mathbf{v}}^H \quad (16)$$

Where: $\tilde{\mathbf{v}}$ is the principal eigenvector of $\hat{\mathbf{R}}_{SS}$ and the desired estimate of the signal vector. In this case the beamformer weight vector is given by:

$$\tilde{\mathbf{w}} = \hat{\mathbf{R}}_{IN}^{-1} \tilde{\mathbf{v}} \quad (17)$$

V. NUMERICAL RESULTS

In this section we examine the behaviour and performance of the proposed method. As the performance metric, Euclidean norm of a projection of the solution onto the vector considered as an “ideal” solution is used. The other metric is the gain in output SINR as compared to one antenna element with a matched filter. An example with the uplink of WCDMA [8] with long scrambling codes is used. A central cell with 5 high data rate (SF=8) and 10 low data rate (SF=256) interfering users is modeled whereas other cells interfering power is modeled as space-time AWGN. Moreover perfect power control and perfect channel estimation is assumed. Eight element uniform linear array with $\lambda/2$ spacing is applied. The radio channel is a local scattering model with azimuthal spread $\sigma_{AZ} = 2^\circ$ and number of multipaths given by Poisson distribution with mean =15. For comparison purposes an “ideal” solution is obtained by removing the desired signal component from the sample correlation matrix of the matched filter outputs.

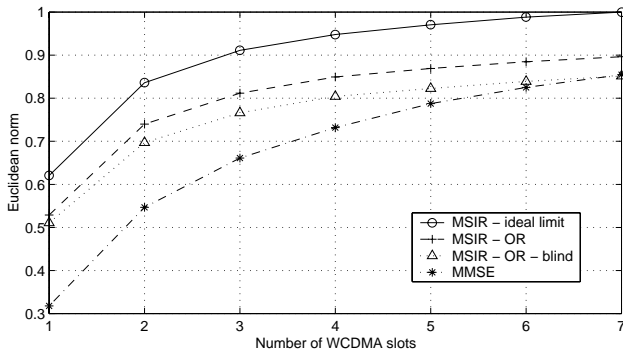


Figure 5: Euclidean norm of a projection of the solution onto a vector considered as an “ideal” solution.

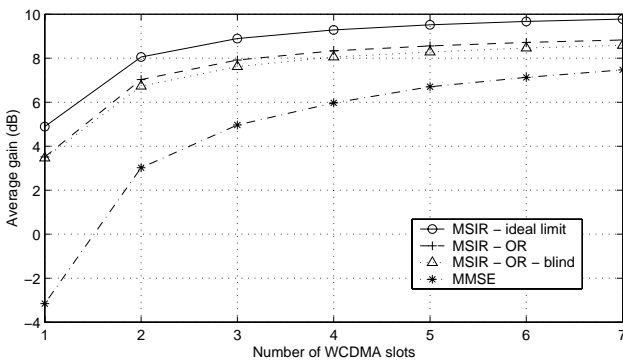


Figure 6: Gain in SINR (dB) as compared to 1 antenna element.

As can be seen from figure 5 the orthogonal re-spread (OR) with the MSIR criterion trails just behind the “ideal” solution and outperforms the MMSE. When translated into SINR gain (figure 6) is about 1dB behind the “ideal” solution. This also applies to the blind orthogonal re-spread method (MSIR -OR - blind). The advantage of the OR over the MMSE is especially evident in the case where the correlation matrices are estimated over very few slots. In

fact when the correlation matrix is estimated over 1 slot only, the 8-element array with the MMSE performs worse than 1 antenna element receiver, which is caused by the desired signal cancellation. The strength of the OR comes from an absence of the desired signal in the matrix of the MSIR solution (4). Consequently, the signal cancellation effect is prevented albeit the estimation errors. The improvement is evident nevertheless $\hat{\mathbf{R}}_{IN}$ is not a statistically consistent estimate of \mathbf{R}_{IN} , which is due to the different MAI realisation when the signal is de-spread with different OVFS code. However the penalty is only ~1dB and is consistent as seen in figure 6.

VI. CONCLUSIONS

The orthogonal re-spread (OR) method possesses the efficacy and simplicity making it very attractive for practical applications. Although this paper introduces the OR method with a simple receiver architecture, the OR is also applicable with more advanced architectures e.g. joint space-time equalizers. The OR method is applicable to DS-CDMA systems, where orthogonal codes used for signal spreading are masked by cell specific scrambling codes as with WCDMA. In particular the OR method suits best in case of long scrambling codes where averaging over different MAI realisations is possible.

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